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Pulse modulator and method for pulse modulation

The invention relates to a pulse modulator for conversion of a complex input signal to a pulsed signal, and to a method for pulse modulation of a complex input signal.

Digital/analog converters may be used to convert a digital input signal to an analog signal. However, these modules are expensive and require a relatively large amount of electrical power. A number of supply voltages are frequently required. A further disadvantage is that digital/analog converters are difficult to integrate with the digital electronics and thus restrict miniaturization.

Digital/analog converters are thus being replaced by digital pulse modulators, such as sigma-delta converters, in many applications. A conventional sigma-delta modulator has an integrator, which integrates the difference signal between the input signal and a fed-back quantized signal, as well as a quantizer, which quantizes the integrated signal. A quantized pulsed signal can then be tapped off at the output of the quantizer, and is fed back as a feedback signal to the input of the sigma-delta converter. Sigma-delta modulators are distinguished by a typical noise characteristic, with the quantization noise being shifted from the low-frequency range in the vicinity of $\omega=0$ towards higher frequencies. The noise which occurs in the region of higher frequencies can then be suppressed with the aid of a downstream low-pass filter. Sigma-delta converters can be implemented at low cost, and can be integrated with the digital electronics. However, for some applications, it would be advantageous to be able to keep the quantization noise in higher frequencies low.

One object of the invention is thus to provide a pulse

modulator as well as a method for pulse modulation, in which the spectral distribution of the quantization noise can be flexibly adapted.

This object of the invention is achieved by a pulse modulator for conversion of a complex input signal to a pulsed signal as claimed in claim 1, by a drive circuit as claimed in claim 16, by a frequency generator as claimed in claim 19, and by a method for pulse modulation of a complex input signal as claimed in claim 21. Claim 31 relates to a computer program product for carrying out the method according to the invention.

The pulse modulator according to the invention for conversion of a complex input signal to a pulsed signal has a subtraction stage which produces a control error signal from the difference between the complex input signal and a feedback signal. The pulse modulator furthermore has a signal conversion stage which converts the control error signal to a control signal. The control signal is multiplied by a complex mixing signal, oscillating at the frequency ω_0, in a first multiplication stage, thus producing at least one of a real part and imaginary part of a control signal which has been up-mixed by ω_0. The pulse modulator furthermore has a quantization stage, which quantizes at least one of the real part and imaginary part of the control signal which has been up-mixed by ω_0 and thus produces the pulsed signal, as well as a feedback unit, which uses the pulsed signal to produce the feedback signal for the subtraction stage.

The method of operation of the pulse modulator according to the invention, which represents an advantageous modification of a conventional sigma-delta converter, will be explained in the following text for the example of an input signal that is kept constant, without any restriction to generality. The subtraction

stage and the signal conversion stage convert this input signal to a control signal, which likewise varies only slightly in time. In contrast to conventional sigma-delta converters, this control signal is, however, now multiplied by the first multiplication stage by a complex mixing signal at the frequency ω_0, in order in this way to produce a control signal up-mixed to the frequency ω_0. The real part or the imaginary part of this control signal oscillating at the frequency ω_0 is then quantized by the quantization stage, thus resulting in a real pulsed signal with a dominant frequency component at the frequency ω_0 at the output of the quantization stage. This real pulsed signal, together with the aid of positive or negative pulses, simulates a sinusoidal signal at the frequency ω_0. This pulsed signal at the same time represents the point of origin for the calculation of the feedback signal, which is fed back to the subtraction stage where it is subtracted from the input signal, in order to determine the control error.

In order to produce the pulsed signal, it is not absolutely essential to calculate both the real part and the imaginary part of the control signal up-mixed by ω_0. If the intention is to derive the pulsed signal from the real part of the up-mixed control signal, then the imaginary part of the up-mixed control signal need not necessarily be produced.

The major advantage of the pulse modulator according to the invention over conventional sigma-delta modulators is that the range of low quantization noise is shifted from the low-frequency range in the vicinity of $\omega=0$ toward the operating frequency ω_0. This is achieved by complex up-mixing of the control signal in the first multiplication stage. This results in a pulsed signal which actually has a low noise level in the relevant spectral range around ω_0.

The starting point for understanding of the noise characteristic is that the signal conversion stage which, for example, may be formed by an integrator, has a low-pass characteristic. This means that relatively high-frequency components are partially suppressed by the signal conversion stage. In conventional sigma-delta converters, this suppression of the higher-frequency components in the control loop causes a rise in the quantization noise at these higher frequencies. In contrast, the quantization noise in the low-frequency range is low. In the case of the pulse modulator according to the invention, the control signal which can be tapped off at the output of the signal conversion stage is up-mixed to the frequency ω_0 by multiplication by the complex mixing signal at the frequency ω_0. The range of low quantization noise is thus also shifted from the frequency $\omega=0$ toward the mixing frequency ω_0, even though the signal conversion stage on the input side is still processing a signal which has not been up-mixed. This results in a pulsed signal with a noise level which is low in the vicinity of ω_0.

The pulse modulator according to the invention can be implemented at low cost, requires relatively little electrical power, and can easily be integrated together with the digital electronics.

It is advantageous for the pulse modulator to have an in-phase signal path for processing of the real part of the input signal, as well as a quadrature signal path for processing of the imaginary part of the input signal. It is also advantageous for the control error signal, the control signal and the feedback signal each to be complex signals, which each have a real signal component as well as an imaginary signal component. In order to ensure that the real pulsed signal reflects the real part or the imaginary part of the control signal up-mixed by ω_0 in the correct phase, the

subtraction stage, the signal conversion stage, the first multiplication stage and the feedback unit are complex signal processing units which each have an in-phase signal path and a quadrature signal path. However, only the real part (or else the imaginary part) of the output signal from the first multiplication stage is required in order to derive the real pulsed signal from it with the aid of the quantization stage. The quantization stage may thus be a real processing stage. In fact, the real pulsed signal is then once again converted to a complex feedback signal in the feedback unit. This design of the pulse modulator makes it possible to synthesize a real pulsed signal, which reproduces a harmonic oscillation at the frequency ω_0 with low phase and amplitude noise, with the correct phase.

According to one advantageous embodiment of the invention, the signal conversion stage has an integrator stage which integrates the control error signal and produces an integrated signal as the control signal. Integration of the control error signal makes it possible to continuously slave the (complex) integrated signal to the complex input signal. Since an integrator stage has a low-pass filter characteristic, this results at the output of the integrator stage in a control signal with a reduced noise level in the region around ω_0. If this control signal is then up-mixed by the first multiplication stage, and is then quantized, this results in a pulsed signal with the desired noise characteristic.

It is advantageous for the integrator stage to have a first integrator for the in-phase signal path and a second integrator for the quadrature signal path, with the first integrator integrating the real part of the control error signal, and with the second integrator integrating the imaginary part of the control error signal. A complex integrator stage for the complex

control error signal can in this way be produced with the aid of two separate integrators.

It is advantageous for the signal conversion stage to have an amplifier stage. The gain factor is in this case chosen such that the quantizer receives the correct input signal level.

According to a further advantageous embodiment of the invention, the first multiplication stage has a first multiplier for the in-phase signal path and a second multiplier for the quadrature signal path. The first multiplier multiplies the real part of the control signal by the real part of the complex mixing signal oscillating at the frequency ω_0, and thus produces a first result signal. The second multiplier multiplies the imaginary part of the control signal by the imaginary part of the complex mixing signal oscillating at the frequency ω_0, and thus produces a second result signal. According to a further advantageous embodiment, the pulse modulator has an adder which adds the first result signal from the first multiplier and the second result signal from the second multiplier to form a sum signal in order to determine the real part of the up-mixed control signal.

If it is assumed that the complex control signal is in the form $R+j\cdot I$, and, by way of example, the complex mixing signal is represented in the form $e^{-j\omega_0 t}$, then the first result signal from the first multiplier becomes $R\cdot\cos(\omega_0 t)$. The second result signal from the second multiplier assumes the form $I\cdot\sin(\omega_0 t)$, and the adder produces the signal $R\cdot\cos(\omega_0 t)+I\cdot\sin(\omega_0 t)$ as the sum signal. However, this signal corresponds precisely to the real part of $(R+j\cdot I)\cdot e^{-j\omega_0 t}$. The real part of the complex multiplication of the control signal and mixing signal can thus be determined by means of the first multiplier, the second multiplier and the adder.

According to one advantageous embodiment of the invention, the sum signal produced by the adder is then quantized by the quantization stage, in order in this way to produce the real pulsed signal.

In this case, it is advantageous for a noise level to be added to the input signal to the quantization stage. The pulse modulator is clocked at a sampling frequency ω_A which must be considerably higher than the mixing frequency ω_0. Certain ratios of ω_0 to ω_A result in relaxation oscillations being formed in the pulse modulator, and these can be seen as additional peaks in the frequency spectrum of the pulsed signal. Since a noise signal is added to the input signal to the quantizer, the result of the quantization process is statistically rounded. This trick makes it possible to prevent the formation of relaxation oscillations.

The quantization stage preferably carries out binary quantization or ternary quantization of its respective input signal. In the case of binary quantization, the pulsed signal may assume only the values 0 and 1. A pulsed signal is thus produced which contains only positive voltage pulses. A ternary-quantized pulsed signal may assume the values -1, 0, 1. A pulsed signal such as this thus comprises both positive and negative voltage pulses. Ternary quantization is thus carried out whenever a pulsed signal is required with both positive and negative pulses.

The feedback unit preferably has a second multiplication stage, which multiplies the pulsed signal by a complex-conjugate mixing signal oscillating at the frequency ω_0, and thus produces the feedback signal down-mixed by ω_0, for the subtractor. The pulsed signal was produced by quantization of the real part of the up-mixed control signal, and thus has its dominant frequency component at the frequency ω_0. Before the pulsed signal can be used as a feedback signal, it must

therefore be down-mixed again to baseband. For this purpose, the pulsed signal is multiplied by a complex-conjugate mixing signal at the frequency ω_0, in order in this way to obtain a down-mixed complex feedback signal.

The second multiplication stage preferably has a third multiplier for production of the real part of the feedback signal and has a fourth multiplier for production of the imaginary part of the feedback signal, with the third multiplier multiplying the pulsed signal by the real part of the complex-conjugate mixing signal oscillating at the frequency ω_0, and with the fourth multiplier multiplying the pulsed signal by the imaginary part of the complex-conjugate mixing signal at the frequency ω_0. In order to shift that frequency component of the pulsed signal which is at the frequency ω_0 in the correct direction, the multiplication of the pulsed signal by the mixing signal must be carried out in complex form. The pulsed signal $y(t)$ is a real signal, while the complex-conjugate mixing signal can be represented in the form $e^{+j\omega_0 t}$. The complex multiplication thus produces a complex feedback signal with the real part $y(t)\cdot\cos(\omega_0 t)$ and the imaginary part $y(t)\cdot\sin(\omega_0 t)$.

The pulse modulator is preferably operated at a sampling frequency ω_A which is 2 to 1000 times higher than the mixing frequency ω_0. This is necessary in order to satisfy the Nyquist condition for the up-mixed signals.

According to a further advantageous embodiment, the pulse modulator is implemented with the aid of a digital signal processor (DSP). All of the operations which are required for operation of the pulse modulator can be programmed with the aid of signal processing routines.

The drive circuit according to the invention for a micromechanical resonator has at least one pulse modulator of the type described above. The pulsed signal which is produced by the at least one pulse modulator is preferably used for electrostatic oscillation stimulation of the resonator. The pulsed signal which is produced can be connected directly to the stimulation electrodes of the resonator. In this case, it is advantageous for the mixing frequency ω_0 of the pulse modulator to correspond to one resonant frequency of the resonator, because this then ensures effective stimulation of the oscillator.

A frequency generator according to the invention for synthesis of a pulsed signal at a predetermined frequency and with a predetermined phase has at least one pulse modulator of the type described above. The pulse modulator according to the invention can be used to produce a corresponding pulsed signal y(t) at a predetermined frequency and with a predetermined phase. In this case, the phase angle of the pulsed signal that is produced can be predetermined very precisely by means of the ratio of the real part and the imaginary part of the input signal x(t). The pulsed signal which is produced has a low noise level in the vicinity of ω_0.

According to a further advantageous embodiment, the pulse modulator is followed by a bandpass filter, preferably a crystal or ceramic filter. This downstream bandpass filter allows those frequency components which are further away from ω_0 and in which the noise level is high to be filtered out.

The invention and further advantageous details will be explained in more detail in the following text with reference to the drawings, which are in the form of exemplary embodiments and in which:

Figure 1 shows a complex block diagram of the pulse modulator according to the invention;

Figure 2 shows a block diagram of the pulse modulator, showing the in-phase path and the quadrature path separately;

Figure 3 shows a ternary-quantized pulsed signal y(t);

Figure 4 shows a frequency spectrum of the pulsed signal y(t) produced at the output of the quantizer;

Figure 5 shows the frequency spectrum from Figure 4 after filtering by a micromechanical oscillator;

Figure 6 shows a frequency spectrum of a pulsed signal y(t) which has been plotted for a ratio of the mixing frequency to the sampling frequency of $\omega_0/\omega_A = 0.25$;

Figure 7 shows a pulse modulator with statistical rounding;

Figure 8 shows the frequency spectrum from Figure 6 with statistical rounding being carried out; and

Figure 9 shows a block diagram of a two-dimensional pulse modulator.

Figure 1 shows a block diagram of the pulse modulator according to the invention, in complex form. The complex input signal x(t) has a real part and an imaginary part, which are both represented as digital values. The complex feedback signal 2 is subtracted from the complex input signal x(t) in the addition node 1, with the difference between these two complex

signals representing the control error. Furthermore, the (likewise complex) content of the delay element 3 is added to this difference in the addition node 1. The content of the delay element 3 is passed via the signal line 4 to the addition node 1. The delay element 3 together with the signal line 4 forms a complex integrator stage, which integrates the complex control error, that is to say the difference between the input signal and the feedback signal. The integrated signal 5 is amplified by the factor "a" in the amplifier stage 6, and the amplified signal 7 is passed to the first multiplication stage 8, where the amplified signal 7 is multiplied by the complex mixing signal $e^{-j\omega_0 t}$ in order in this way to obtain the signal 9, up-mixed to the frequency ω_0. The block 10 determines the real part of the complex up-mixed signal 9, and the real part 11, obtained in this way, of the up-mixed signal is made available to the quantizer 12.

In the embodiment shown in Figure 1, the quantizer 12 is in the form of a ternary quantizer, which converts the respective input signal to the three possible values -1, 0, +1 of a pulsed signal with the aid of comparators. The quantized pulsed signal y(t) produced in this way can be tapped off at the output of the quantizer 12. The real pulsed signal y(t) is multiplied in the second multiplication stage 13 by the complex-conjugate mixing signal $e^{-j\omega_0 t}$ in order to produce the complex feedback signal 2. The complex feedback signal 2, which is obtained in this way by multiplication of a real number and a complex number, is passed to the addition node 1 at the input to the circuit.

The sequence of functional units illustrated in Figure 1 can be implemented by means of a digital signal processor (DSP) or else by means of hardware that is specifically provided for this purpose. The digital signal processing must in this case be carried out at a sampling frequency ω_A, which is considerably

higher than the frequency ω_0 of the complex mixing signal. For example, 2 to 1000 times the mixing frequency ω_0 may be used as the sampling rate ω_A.

Figure 2 once again shows the pulse modulator illustrated in Figure 1, with the in-phase signal path and the quadrature signal path in this case being shown separately. The upper half of Figure 2 shows the in-phase signal path 14, which processes the real part R of the input signal x(t). The lower half of Figure 2 shows the quadrature signal path 15 for processing of the imaginary part I of the input signal. The real part of the control error is determined in the addition node 16 in the in-phase path as the difference between the real part R of the input signal and the real part 17 of the feedback signal. The integrator value, which has been stored until then in the delay element 18, is added to this control error, and is passed via the signal line 19 to the addition node 16. Together with the signal line 19, the delay element 18 forms an integrator with the transfer function $H(z) = \frac{1}{1 - z^{-1}}$. Addition of the real part of the control error to the previous integrator value results in a new integrator value, which is once again stored in the delay element 18. The integrated signal 20 in the in-phase signal path is scaled by the factor "a" by the amplifier 21, and is passed as the amplified signal 22 to the first multiplier 23. The first multiplier 23 multiplies the real, amplified signal 22 by the real signal $\cos(\omega_0 t)$, that is to say by the real part of $e^{-j\omega_0 t}$. The first multiplier 23 determines the product $R \cdot \cos(\omega_0 t)$, which is supplied as the signal 24 to the adder 25.

The quadrature signal path 15 of the pulse modulator has an addition node 26, in which the difference between the imaginary part I of the input signal and the imaginary part 27 of the feedback signal is calculated. This difference, which corresponds to the

imaginary part of the control error, is added to the previous content of the delay element 28, which is passed to the addition node 26 via the signal line 29. The new value, which is obtained as the sum of the previous value and of the imaginary part of the control error, is written to the delay element 28. Together with the signal line 29, the delay element 28 forms an integrator with the transfer function $H(z) = \frac{1}{1 - z^{-1}}$. The integrated signal 30 from the quadrature signal path is produced at the output of this integrator, and is scaled by the factor "a" by the amplifier 31. The amplified signal 32 obtained in this way in the quadrature signal path is then multiplied by the signal $\sin(\omega_0 t)$ in the second multiplier 33. The product $I \cdot \sin(\omega_0 t)$ obtained in this way is supplied as the signal 34 to the adder 25. The adder 25 adds the signals $R \cdot \cos(\omega_0 t)$ and $I \cdot \sin(\omega_0 t)$ and produces the signal $R \cdot \cos(\omega_0 t) + I \cdot \sin(\omega_0 t)$ as the signal 35 at its output. However, this signal 35 corresponds precisely to the real part of the up-mixed signal, because the complex multiplication of $x(t)$ and $e^{-j\omega_0 t}$ gives:

$$x(t) \cdot e^{-j\omega_0 t} =$$
$$= (R + j \cdot I) \cdot (\cos(\omega_0 t) - j \cdot \sin(\omega_0 t)) =$$
$$= [R \cdot \cos(\omega_0 t) + I \cdot \sin(\omega_0 t)] + j \cdot [I \cdot \cos(\omega_0 t) - R \cdot \sin(\omega_0 t)]$$

and the real part of this signal is $R \cdot \cos(\omega_0 t) + I \cdot \sin(\omega_0 t)$. The signal 35 thus represents the real part of the complex up-mixed signal, and to this extent corresponds to the signal 11 illustrated in Figure 1.

The digital real signal 35 is passed to the quantizer 36, which converts this input signal to the quantized pulsed signal $y(t)$. The three-stage (ternary) quantizer shown in the example in Figure 1 and Figure 2 quantizes the input signal on the basis $y(t) \in \{-1;\ 0;\ +1\}$. For this purpose, the quantizer 36 has comparators, which

compare the signal level of the signal 35 continuously with predetermined threshold values. Depending on the result of these comparisons, the output signal y(t) is in each case assigned one of the values -1; 0; +1 as the current signal value. Instead of the three-stage (ternary) quantization, any other desired quantizations may be used depending on the purpose, for example two-stage (binary) or multiple-stage quantizations.

The real part 17 and the imaginary part 27 of the complex feedback signal are derived from the quantized pulsed signal y(t). For this purpose, the pulsed signal y(t) is multiplied by the complex-conjugate mixing signal $e^{+j\omega_0 t}$:

$$y(t)\cdot e^{j\omega_0 t}=y(t)\cdot\cos(\omega_0 t)+j\cdot y(t)\cdot\sin(\omega_0 t)$$

The real part $y(t)\cdot\cos(\omega_0 t)$ of the complex feedback signal is produced by the third multiplier 37, which multiplies the pulsed signal y(t) by $\cos(\omega_0 t)$. The real part 17 of the feedback signal is thus produced at the output of the third multiplier 37, and is fed back to the addition node 16. In order to produce the imaginary part $y(t)\cdot\sin(\omega_0 t)$ of the complex feedback signal, the pulsed signal y(t) is multiplied by $\sin(\omega_0 t)$ in the fourth multiplier 38. The imaginary part 27 of the feedback signal is produced at the output of the fourth multiplier 38, and is fed back to the addition node 26.

In the exemplary embodiments shown in Figures 1 and 2, integrators are provided on the input side, which integrate the control error between the input signal and the feedback signal, and thus produce an integrated signal. The transfer function H(z) of an integrator can be written as $H(z)=\frac{1}{1-z^{-1}}$. Other signal conversion stages with other transfer functions H(z) may also be used on the input side, instead of the integrators. For example, higher-order transfer functions H(z) could be

used in which case, however:

$$\lim_{z \to 1} H(z) = \infty$$

The transfer function H(z) should thus tend to infinity for the situation in which the frequency ω tends to the value zero ($z \to 1$). The additional free parameters of H(z) may be used to optimize specific characteristics of the modulator (for example the signal-to-noise ratio) or of the overall system.

Figure 3 shows the waveform of the pulsed signal y(t) which can be tapped off at the output of the quantizer for the situation of ternary quantization with $y(t) \in \{-1; 0; +1\}$, which was determined with the aid of a computer simulation. In this case, the real part R of the complex input signal was set to 0.3, while the imaginary part I of the input signal was set to be equal to zero. The input signal x(t) is thus constant, and does not vary as a function of time. The sampling frequency ω_A is five times as great as the mixing frequency $\omega_0/\omega_A = 0.2$. The clock pulses at the sampling frequency ω_A are shown on the abscissa, and are numbered successively from 5000 to 5100. During each clock cycle, the pulsed signal y(t) assumes one of the three possible values -1; 0; +1. The respective value of y(t) during one specific clock cycle at the sampling frequency is plotted in the direction of the ordinate.

If a spectral analysis (FFT) is carried out on the pulsed signal illustrated in Figure 3, this results in the spectrum shown in Figure 4. The frequency of the respective spectral components is shown in arbitrary FFT units on the abscissa, while the signal intensity is plotted in dB in the direction of the ordinate. A peak can be seen in the spectral distribution at the frequency ω_0. It can also be seen that the noise level in the vicinity of the frequency ω_0 is considerably

less than in the remaining part of the spectrum. In a conventional sigma-delta modulator, the noise level would in contrast be reduced considerably at low frequencies, that is to say in the vicinity of the frequency ω_0. In the case of the pulse modulator according to the invention, the integrated and amplified signal is up-mixed to the mixing frequency ω_0 by means of a complex multiplication. In consequence, the spectral range in which the noise is reduced is also shifted toward the mixing frequency ω_0, thus resulting in the noise characteristic illustrated in Figure 4.

The pulse modulator according to the invention can be used for digital synthesis of a pulsed signal, in which case the main spectral component of the pulsed signal can be predetermined by the mixing frequency ω_0. The phase angle of the pulsed signal that is produced can be set exactly by the ratio of the real part to the imaginary part of the input signal, and this results in a pulsed signal whose phase is stable. When using the pulse modulator according to the invention for frequency synthesis, the pulsed signal y(t) should be filtered by means of an electrical bandpass filter, whose pass band is centered around the frequency ω_0. This bandpass filter which, for example, may be in the form of a crystal or ceramic filter, makes it possible to suppress spectral ranges further away from ω_0, in which the noise level is undesirably high. A bandpass filter such as this makes it possible to significantly improve the signal-to-noise ratio.

The pulse modulator according to the invention is suitable, inter alia, for stimulation of electromechanical oscillators to carry out harmonic oscillations. In particular, the electrostatic forces which are required for oscillation stimulation can be produced by means of a ternary-quantized pulsed signal which is applied to the stimulation electrodes of a

micromechanical resonator. The frequency ω_0 of the pulsed signal y(t) is in this case preferably chosen to be equal to the resonant frequency of the micromechanical oscillator. If the pulsed signal as illustrated in Figure 3 and Figure 4 is used for harmonic stimulation of a high Q-factor oscillator (for example with a Q-factor of 10^4), whose resonant frequency corresponds to the stimulation frequency ω_0, then the majority of the quantization noise is filtered out by the oscillator itself. In particular, the quantization noise in spectral ranges further away from the resonant frequency ω_0 is suppressed by the oscillator itself. The filtered spectrum obtained in this way is shown in Figure 5.

Specific ratios of the frequencies ω_0/ω_A exist for which the noise-like quantization product in y(t) is converted to a series of more or less periodic functions. As one example of this, Figure 6 shows a frequency spectrum which was obtained for the ratio $\omega_0/\omega_A = 0.25$. A range of spectral lines 39, 40, 41, etc. can be seen in addition to the peak at the frequency ω_0. The reason for the creation of these spectral lines is that the quantizer is a highly non-linear element in the control loop, because this stimulates relaxation oscillations in the control loop with certain frequency ratios. This control loop response is known from conventional delta-sigma converters.

In order to prevent the creation of relaxation oscillations, the central linearity of the quantizer can be improved by adding a noise signal to the input signal to the quantizer. A spectrally uniformly distributed noise signal is preferably used for this purpose. Figure 7 shows the block diagram of a correspondingly modified pulse modulator. In comparison to the block diagram shown in Figure 2, the pulse modulator shown in Figure 7 additionally has a noise

generator 42, which produces a noise signal 43. In addition, the integrators which are shown in Figure 2 are illustrated in a generalized form as signal conversion stages 44, 45 with the transfer function H(z). Otherwise, the assemblies shown in Figure 7 correspond to the elements of the block diagram in Figure 2. The noise signal 43 is supplied to the adder 25, where it is added to the signals 24 and 34. The signal 35 at the input of the quantizer 36 therefore has a noise signal superimposed on it, and, in the end, this leads to statistical rounding in the quantization process. Figure 8 shows the frequency spectrum of a pulsed signal y(t) which was produced with the aid of a pulse modulator modified as shown in Figure 7. Although the frequency ratio ω_0/ω_A is once again equal to 0.25, no relaxation oscillations are formed.

The pulse modulator according to the invention can be used in particular for electrostatic stimulation of micromechanical oscillators. For this purpose, by way of example, a ternary-quantized pulsed signal of the type shown in Figure 3 can be connected to the stimulation electrodes of a micromechanical resonator. The pulsed signal shown in Figure 3 represents a sinusoidal signal at the frequency ω_0. A pulsed signal such as this can thus be used to stimulate a micromechanical resonator to carry out harmonic oscillations at the frequency ω_0, to be precise in particular when the frequency ω_0 of the pulsed signal corresponds at least approximately to the resonant frequency of the oscillator.

Resonators which can oscillate in two mutually perpendicular directions y_1 and y_2 are used in rotation rate sensors and Coriolis gyros. The two-dimensional pulse modulator shown in Figure 9 may be used for electrostatic stimulation of a resonator with two degrees of freedom. The two-dimensional pulse modulator has a first pulse modulator 46, which produces the

pulsed signal $y_1(t)$ from the complex input signal R_1, I_1, and this pulsed signal is used to stimulate the resonator in the y_1 direction. The pulsed signal $y_2(t)$ is produced from the complex input signal R_2, I_2 by the second pulse modulator 47, and this pulsed signal is used to stimulate the oscillator to oscillate in the y_2 direction. Both the first pulse modulator 46 and the second pulse modulator 47 are in the form of a pulse modulator with statistical rounding as shown in Figure 7. A description of the design and method of operation of the first and of the second pulse modulator 46, 47 can therefore be found in the description of the figures relating to Figures 2 and 7. However, the two-dimensional pulse modulator shown in Figure 9 has one 2D quantizer 48 which is shared by the two channels and converts the signal 49 of the first pulse modulator 46 to the quantized pulsed signal $y_1(t)$, and transforms the signal 50 of the second pulse modulator 47 to the quantized pulsed signal $y_2(t)$. The use of a 2D quantizer 48 which is shared by the two channels makes it possible during the quantization of the signals 49, 50 to take into account additional conditions which are advantageous for operation of the micromechanical sensor. One such additional condition, by way of example, is that in each case only one of the channels may produce pulses other than zero. Another feasible additional condition is that only one of the output signals $y_1(t)$, $y_2(t)$ may change in each case at any given time. Additional conditions such as these may be worthwhile when the displacement currents which are applied to the electrodes of a double resonator are measured in sum form, in order to make it possible to deduce the deflection of the oscillator. The additional conditions make it possible to unambiguously associate a displacement current with one specific electrode. This makes it possible to carry out signal separation between the signals caused by the y_1 deflection and the y_2 deflection of the oscillator.